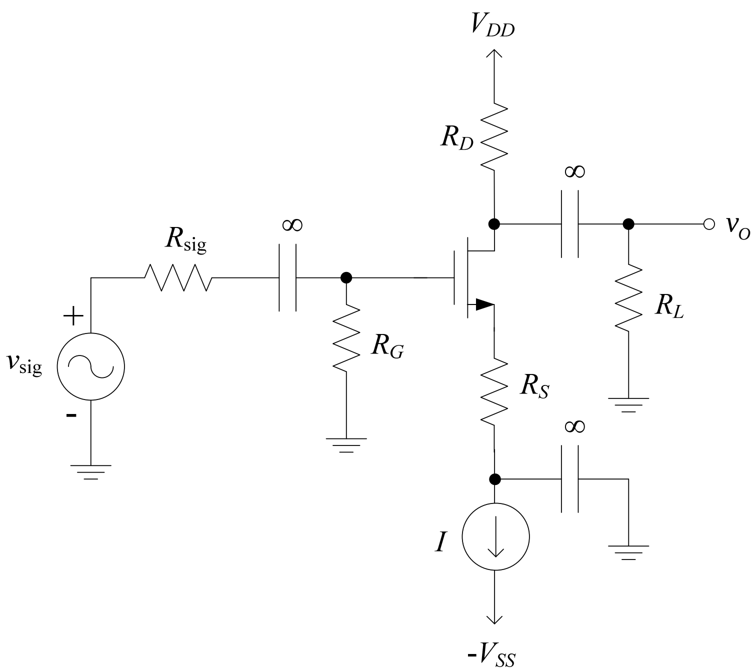
Lecture 32: Common Source Amplifier with Source Degeneration.

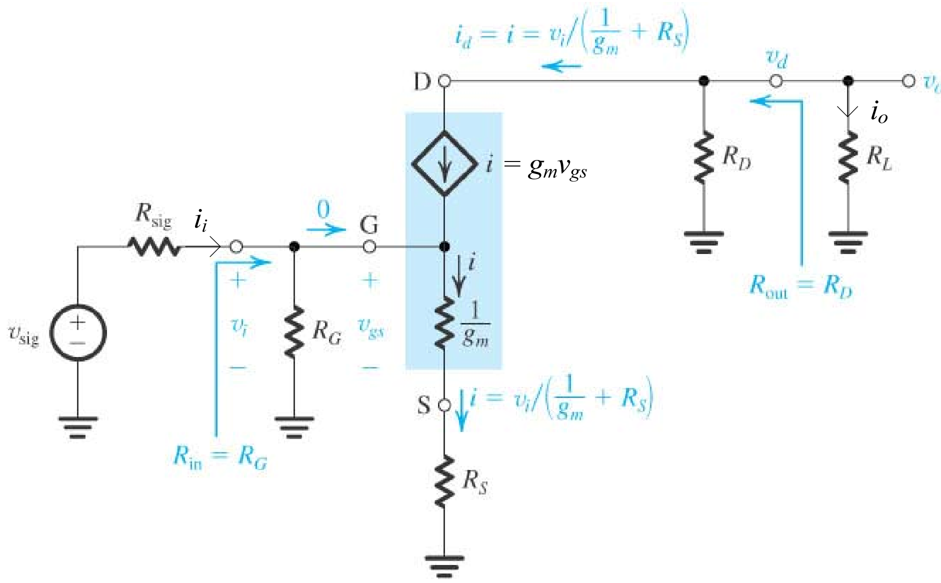
The small-signal amplification performance of the CS amplifier discussed in the previous lecture can be improved by including a series resistance in the source circuit. (This is very similar – if not identical – to the effect of adding emitter degeneration to the BJT CE amplifier.) This so-called CS amplifier with source degeneration circuit is shown in Fig. 4.44(a).



(Fig. 4.44a)

We have a choice of small-signal models to use for the MOSFET. A T model will simplify the analysis, on one hand, by allowing us to incorporate the effects of R_S by simply adding this value to $1/g_m$ in the small-signal model, if we ignore r_o.

This small-signal circuit is shown in Fig. 4.44(b).



(Fig. 4.44b)

On the other hand, using the T model makes the analysis more difficult when r_o is included. (The hybrid π model is better at easily including the effects of r_o.) However, r_o in the MOSFET amplifier is large so we can reasonably ignore its effects for now in the expectation of making the analysis more tractable.

Small-Signal Amplifier Characteristics

We'll now calculate the following small-signal quantities for this MOSFET common source amplifier with source degeneration: R_{in}, A_v, G_v, G_i, and R_{out}.

- Input resistance, R_{in}. Referring to the small-signal equivalent circuit above in Fig. 4.44(b), with $i_g = 0$, then

$$R_{\text{in}} = R_G \tag{4.84),(1}$$

- Partial small-signal voltage gain, A_v. We see at the output side of the small-signal circuit in Fig. 4.44(b)

$$v_o = -g_m v_{gs} \left(R_D \parallel R_L \right) \tag{2}$$

which is the same result (ignoring r_o) as we found for the CS amplifier without source generation. At the gate, however, we find through voltage division that

$$v_{gs} = \frac{1/g_m}{1/g_m + R_S} v_i = \frac{v_i}{1 + g_m R_S} \tag{4.86),(3}$$

This is a different result than for the CS amplifier in that v_{gs} is only a fraction of v_i here, whereas $v_{gs} = v_i$ without R_S.

Substituting (3) into (2), gives the partial small-signal AC voltage gain to be

$$A_v \equiv \frac{v_o}{v_i} = \frac{-g_m \left(R_D \parallel R_L \right)}{1 + g_m R_S} \tag{4.88),(4}$$

- Overall small-signal voltage gain, G_v. As we did in the previous lecture, we can derive an expression for G_v in terms of A_v. By definition,

$$G_v \equiv \frac{v_o}{v_{\text{sig}}} = \frac{v_i}{v_{\text{sig}}} \underbrace{\frac{v_o}{v_i}}_{=A_v} = \frac{v_i}{v_{\text{sig}}} A_v \tag{5}$$

Applying voltage division at the input of the small-signal equivalent circuit in Fig. 4.44(b),

$$v_i = \frac{R_{\text{in}}}{R_{\text{in}} + R_{\text{sig}}} v_{\text{sig}} \underbrace{=}_{(1)} \frac{R_G}{R_G + R_{\text{sig}}} v_{\text{sig}} \tag{6}$$

Substituting (6) into (5) we the overall small-signal AC voltage gain for this CS amplifier with source degeneration to be

$$G_v = \frac{-R_G}{R_G + R_{\text{sig}}} \frac{g_m \left(R_D \,\|\, R_L\right)}{1 + g_m R_S} \qquad (4.90),(7)$$

- <u>Overall small-signal current gain, G_i</u>. Using current division at the output in the small-signal model above in Fig. 4.44(b)

$$i_o = \frac{-R_D}{R_D + R_L} g_m v_{gs} \tag{8}$$

while at the input,

$$i_i = \frac{v_i}{R_G} \underbrace{=}_{(3)} \frac{1 + g_m R_S}{R_G} v_{gs} \tag{9}$$

Substituting (9) into (8) we find that the overall small-signal AC current gain is

$$G_i \equiv \frac{i_o}{i_i} = \frac{-g_m R_D}{R_D + R_L} \frac{R_G}{1 + g_m R_S} \tag{10}$$

- <u>Output resistance, R_{out}</u>. From the small-signal circuit in Fig. 4.44(b) with $v_{\text{sig}} = 0$ then i must be zero leading to

$$R_{\text{out}} = R_D \tag{11}$$

Discussion

Adding R_S has a number of effects on the CS amplifier. (Notice, though, that it doesn’t affect the input and output resistances.)

First, observe from (3)

$$v_{gs} = \frac{v_i}{1+g_m R_S} \tag{3}$$

that we can employ R_S as a tool to lower v_{gs} relative to v_i and lessen the effects of nonlinear distortion.

This R_S also has the effect of lowering the small-signal voltage gain, which we can directly see from (7).

A major benefit, though, of using R_S is that the small-signal voltage (and current) gain can be made much less dependent on the MOSFET device characteristics. (We saw a similar effect in the CE BJT amplifier with emitter degeneration.)

We can see this here for the MOSFET CS amplifier using (7)

$$G_v = \frac{-R_G}{R_G + R_{\text{sig}}} \frac{g_m \left(R_D \parallel R_L\right)}{1+g_m R_S} \tag{7}$$

The key factor in this expression is the second one. In the case that $g_m R_S \gg 1$ then

$$G_v \approx \frac{-R_G}{R_G + R_{\text{sig}}} \frac{R_D \parallel R_L}{R_S} \tag{12}$$

which is no longer dependent on g_m.

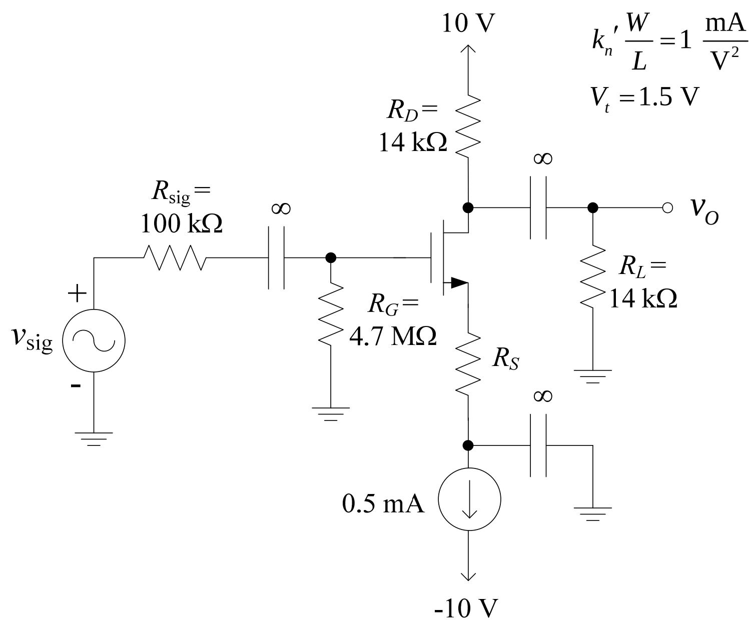
Conversely, without R_S in the circuit ($R_S = 0$), we see from (7) that $G_v \propto g_m$ and is directly dependent on the physical properties of the transistor (and the biasing) because

$$g_m \equiv \frac{i_d}{v_{gs}} = k_n' \frac{W}{L}\left(V_{GS} - V_t\right) \tag{4.61),(13}$$

in the case of an NMOS device.

The "price" we pay for this desirable behavior in (12) – where G_v is not dependent on g_m – is a reduced value for G_v. This G_v is largest when $R_S = 0$, as can be seen from (7).

Example N32.1 (based on text exercises 4.32 and 4.33). Compute the small-signal voltage gain for the circuit below with $R_S = 0$, $k_n' W/L = 1$ mA/V^2, and V_t = 1.5 V. For a 0.4-V_{pp} sinusoidal input voltage, what is the amplitude of the output signal?



For the DC analysis, we see that $V_G = 0$ and $I_D = I_S = 0.5$ mA. (Why is $V_G = 0$?) Consequently,

$$V_D = 10 - R_D I_D = 10 - 14\text{k} \cdot 0.5\text{m} = 3 \text{ V}$$

Assuming MOSFET operation in the saturation mode

$$I_D = \frac{1}{2} k_n' \frac{W}{L} \left(V_{GS} - V_t\right)^2$$

such that $$0.5 \text{ mA} = \frac{1}{2} 1 \times 10^{-3} \left(V_{GS} - 1.5\right)^2$$

or $$V_{GS} - 1.5 = \pm 1 \quad \Rightarrow \quad V_{GS} = 2.5 \text{ V or } 0.5 \text{ V}$$

Therefore, $$V_S = -2.5 \text{ V}$$

for operation in the saturation mode.

For the AC analysis, from (13)

$$g_m = 10^{-3}\left(2.5 - 1.5\right) = 1 \text{ mS}$$

Using this result in (7) with $R_S = 0$ gives

$$G_v = \frac{-4.7\text{M}}{4.7\text{M}+100\text{k}} 10^{-3}\left(14\text{k} \,\|\, 14\text{k}\right) = -6.85 \ \frac{\text{V}}{\text{V}}$$

For an input sinusoid with 0.4-V_{pp} amplitude, then

$$V_o = G_v \cdot V_{\text{sig}} = 6.85 \cdot 0.4 \ \text{V}_{\text{pp}} = 2.74 \ \text{V}_{\text{pp}}$$

Will the MOSFET remain in the saturation mode for the entire cycle of this output voltage? For operation in the saturation mode, $v_{DG} = v_D > V_t = 1.5$ V. On the negative swing of the output voltage,

$$v_D\big|_{\text{min}} = V_D - \frac{v_{o,\text{pp}}}{2} = 3 - \frac{2.74}{2} = 1.63 \ \text{V}$$

which is greater than V_t, so the MOSFET will not leave the saturation mode on the negative swings of the output voltage. On the positive swings,

$$v_D\big|_{\text{max}} = V_D + \frac{v_{o,\text{pp}}}{2} = 3 + \frac{2.74}{2} = 4.37 \ \text{V}$$

which is less than $V_{DD} = 10$ V so the MOSFET will not cutoff and leave the saturation mode.

(Interestingly, the MOSFET **does** leave the saturation mode on the negative swings for $R_D = R_L = 15$ kΩ, as used in the text exercises 4.32 and 4.33.)

Lastly, imagine that for some reason the input voltage is increased by a factor of 3 (to 1.2 V_{pp}). What value of R_S can be used to keep the output voltage unchanged?

From (7), we can choose R_S so that the so-called feedback factor $1+g_m R_S$ equals 3. The output voltage amplitude will then be unchanged with this increased input voltage.

Hence, for

$$1+g_m R_S = 3 \quad \Rightarrow \quad R_S = \frac{3-1}{g_m} = \frac{2}{10^{-3}} = 2 \text{ k}\Omega.$$

With $R_S = 2$ kΩ the new overall small-signal AC voltage gain is from (7)

$$G_v = \frac{-6.85}{1+g_m R_S} = \frac{-6.85}{3} = -2.28 \ \frac{\text{V}}{\text{V}}$$

The overall small-signal voltage gain has gone down, but the amplitude of the output voltage has stayed the same since the input voltage amplitude was increased.